TUNING THE ANGULAR SPEED CONTROLLER OF THE REACTIVE FLYWHEEL OF THE NANOSATELLITE ORIENTATION SYSTEM

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The paper describes reactive flywheels based on brushless magnetoelectric motors for controlling the spatial orientation of satellites, developed at the Institute of Electrodynamics of the National Academy of Sciences of Ukraine. The main attention is paid to the problem of implementing a rotor angular speed and position sensor, provided that it can be integrated into the motor design. The influence of the number of feedback signal pulses per one revolution of the motor shaft on the tuning of the reactive flywheel angular speed controller is studied under the condition that the ripple range at the controller output signal is limited at a given level. The dependences of the filter time constant and the controller gains on the value of the reactive flywheel angular speed and on the number of feedback signal pulses per revolution are obtained. Ref. 8, fig. 5, table.
Key words: nanosatellite, reactive flywheel, brushless magnetoelectric motor, control system, angular speed sensor, angular speed controller.

Introduction. The development of the technology of nano- and microsatellites is primarily associated with the tendency to reduce the cost of spacecraft designed to solve specific applied problems in space, as well as the need to operate satellites as part of a group, which expands the range of tasks that cannot be solved by a single spacecraft. Today, there is a conditional classification of such satellites, according to which devices weighing from 0.1 kg to one kilogram are assigned to the class of picosatellites, from 1 kg to 10 kg - to nanosatellites and from 10 kg to 100 kg - to microsatellites [1]. The variety of forms of vehicles and the possibility of simultaneously launching several satellites using one launch vehicle predetermined their standardization in terms of geometric dimensions. This is how the CubeSat standard appeared, according to which the satellite has a size of $10\times10\times10$ cm and a weight of 600 grams to 1.33 kg. A satellite may consist of several of these units, therefore, according to the classification by mass, it can correspond to any of the types listed above.

Like any spacecraft, CubeSat satellites are structurally composed of one or more onboard computers, active and passive orientation systems, power sources - solar panels and batteries, communication systems with a control center, payload, and so on. Placing all these systems in a limited volume is a complex task that requires the developer to constantly search for compromise solutions between cost, control quality and weight and size indexes. Particularly acute is the problem of ensuring all consumers with electricity, the largest of which and the most expensive is an active attitude control system constructed using reactive flywheels. The use of reactive flywheels to control the angular position of the satellite in the best way allows you to perform the functions of an attitude control system, namely: accurate orientation of the spacecraft in space, reliable stabilization of its position and high speed when compensating for unwanted disturbances acting on the device.

The operation of the satellite orientation system in space using reactive flywheels is based on the use of the fundamental law of conservation of momentum in a closed system of bodies. Since negligible external forces act on a spacecraft in outer space, any interaction of moving bodies associated with the satellite body leads to a change in its spatial position. The principle of orientation using flywheels is that when an electromagnetic torque is applied to the flywheel rotor, its stator is affected by an equal torque of the opposite sign, called the reactive moment. Since the flywheel stator is rigidly fixed in the satellite body, the entire spacecraft is exposed to this effect and makes a turn around the flywheel axis in the opposite direction from the direction of the flywheel motor torque. Depending on the tasks solved by the space mission, there can be from one to four

flywheels on board [2, 3], which operate during the entire flight and are, as a rule, the main consumers of electricity. Therefore, in the vast majority of cases, for the implementation of reactive flywheels, the most economical and small-sized among all electrical machines are used brushless electric motors with permanent magnets [4–6].

In [4], an overview of the implementation variants for reactive flywheels is presented, as well as the principles for development of the structure proposed by the authors. When designing reactive flywheels, a number of important factors must be taken into account, but this paper will consider only some of the features of tuning a control system.

The purpose of the paper is to present the experience of developing reactive flywheels for satellite control, obtained at the Institute of Electrodynamics of the National Academy of Sciences of Ukraine, as well as to study the features of tuning the angular speed controllers depending on the characteristics of the motor mechanical coordinates sensor.

The main material and research results. The authors of the paper already have successful experience in creating a reactive flywheel (Fig. 1 *a*) for the PolyITAN-2-SAU nanosatellite, developed at the National Technical University of Ukraine "KPI", which was launched into the Earth's orbit in May 2017 and successfully completed the entire mission period [7]. The satellite was equipped with a flywheel for a single-axis attitude control system. In the flywheel drive, a brushless magnetoelectric motor (BMM) with an external rotor and an integrated control system was used [4]. In continuation of this work, flywheels were created for the triaxial layout of the attitude control system (Fig. 1 *b*).

Table shows the main characteristics of two types of flywheels (Fig. 1).

The BMM-based reactive flywheel is a space-use device that has specific require-ments. Some of the features of its implementation will be discussed below.

The reactive flywheel is subject to increased requirements to achieve minimum weight and size indexes, economy of operation and high reliability. In addition, it is necessary

a   **Fig. 1**   *b*

to provide a zero error when processing the input reference signal, which is achieved by implementing an angular speed astatic control system.

The nature of the reactive flywheel operation mode is determined by the relatively large moment of inertia on the motor shaft, while its mechanical load is determined only by the resistance of the bearings, eddy current losses in the windings, and also by the aerodynamic resistance of the rotating rotor. The latter is possible only in terrestrial conditions or when the body of the motor-flywheel

Parameter	Flywheel motor (Fig. 1*a*)	Flywheel motors (Fig. 1*б*)
Kinetic moment	0,0135 Nms	0,09 Nms
Speed range	±6000 rpm	±6000 rpm
Maximum power consumption at 6000 rpm	0,012 W	0,018 W
Maximum allowed power consumption	1 W	2 W
Maximum motor torque	0,05 Nm	0,15 Nm
Mass	0,2 kg	0,42 kg
Overall dimensions	$\Phi 52 \times 42$ mm	$74 \times 74 \times 33$ mm
Supply voltage	2,6...3,6 V	2,6...3,6 V
Operating temperature	-30...+50 °C	-30...+50 °C

is sealed. In space, the motor load is determined only by the bearings resistance torque and eddy current losses in the windings.

Under the conditions of structural minimization of the reactive flywheel hardware, there are special requirements for the implementation of the mechanical coordinate sensor, with the help of which signals are generated corresponding to the angular speed and position of the rotor. To develop such a sensor, one should strive to use existing motor elements, or the additional elements introduced should be install into the motor design, that is, the integration of mechanical coordinate sensor elements into the flywheel design should not lead to an increase in the overall dimensions of the entire device.

One of the variants for constructing a mechanical coordinate sensor that satisfies this principle is the integration of Hall sensors into the active zone of an electric motor to form a sequence of pulses. In the simplest case, when installing discrete Hall sensors, according to the number of phases of the stator winding, you can easily form a sequence of pulses, the number of which per revolution of the flywheel shaft is equal to

$$N = 2\,p\,m\,, \tag{1}$$

where p is the number of pairs of poles; m is the number of motor phases. In this case, the pulse reiteration period is determined by the formula

$$T_N = \frac{2\pi}{N\omega}, \tag{2}$$

where ω is the angular speed of the rotor.

When using analog Hall sensors, the number N of pulses can be increased [8].

Let's study the modes of operation of the reactive flywheel. Under the conditions of a sinusoidal distribution of magnetic induction in the gap, electrical and magnetic symmetry of the motor, as well as neglecting losses in iron, the mathematical model of a brushless motor with a slotless stator and when a surface installation of magnets on the rotor in the coordinate system of the rotor d and q [8] has the form

$$L\,\frac{d\,i_d}{dt} = -R\,i_d + p\,L\,\omega\,i_q + u_d\,;\ L\,\frac{d\,i_q}{dt} = -R\,i_q - p\,L\,\omega\,i_d - k_m\,\omega + u_q\,;\ e = k_{m1}\,\omega\,; \tag{3-5}$$

$$J\,\frac{d\,\omega}{dt} = \Delta M\,;\ \Delta M = M - M_L\,;\ \frac{d\,\theta}{dt} = \omega\,;\ M = 0{,}5\,m\,k_{m1}\,i_q\,, \tag{6-9}$$

where i_d, i_q, u_d, u_q are stator currents and voltages in the rotor coordinate system d and q; R, L are active resistance and inductance of the stator winding; k_{m1} is motor torque coefficient; J is the moment of inertia of the flywheel rotor; ΔM is dynamic torque; M is motor torque; M_L is mechanical load torque; θ is the angle of rotation of the rotor.

The proposed flywheel motor has the following parameters $p = 2$, $m = 3$, $R = 0{,}313\,Ohm$, $L = 2{,}2\cdot10^{-5}\,H$, $J = 1{,}399\cdot10^{-5}\,kg\,m^2$, maximum values of the angular speed and amplitude of the stator EMF are $\omega_{\max} = 628{,}318\,s^{-1}$ and $E_{1\max} = 1{,}732\,V$. The electromagnetic time constant of the stator winding is defined as $T_E = L/R = 7{,}029\cdot10^{-5}\,s$, and the motor torque coefficient k_{m1} depends on the characteristics of the motor core and for the system of equations (3-9) is defined as $k_{m1} = E_{1\max}/\omega_{\max} = 0{,}002757\,Tm^2$.

Considering that the electromagnetic time constant T_E of the stator winding of the BMM is relatively small, and also that the flywheel motor rotates under a small mechanical load, we can neglect the direct component of the stator current, that is, assume $i_d = 0$. In this case, equation (3) is eliminated.

The smallness of the value of the electromagnetic time constant T_E of the stator windings affects the choice of the structure of the semiconductor converter for the formation of stator currents. In this case, when the motor windings are powered by a voltage inverter with pulse-width modulation, current smoothing becomes impossible without increasing the switching frequency.

Then there are two variants for solving the problem. Either introduce additional phase inductances into the converter circuit, which will lead to an increase in weight and size and is an unacceptable solution, or generate stator currents from a voltage inverter without pulse-width control when a smoothed regulated voltage is connected to the inverter input. In the latter case, the structure "BMM - voltage inverter" becomes similar in some properties to a DC collector motor. Neglecting the influence of the electromagnetic time constant T_E of the stator winding due to its smallness, the BMM equations can be written in a simpler form

$$2L\frac{dI}{dt} = -2RI - E + U \; ; \; E = k_{m2}\omega \; ; \; J\frac{d\omega}{dt} = \Delta M \; ; \qquad (10\text{-}12)$$

$$\Delta M = M - M_L \; ; \; \frac{d\theta}{dt} = \omega \; ; \; M = k_{m2}I \; , \qquad (13\text{-}15)$$

where I, E, U are the current and EMF of the stator, as well as the voltage in the DC link of the voltage inverter; k_{m2} is motor torque coefficient, which for the system of equations (10-15) at the maximum value of the motor rectified EMF $E_{\max} = 3\,V$ is defined as $k_{m2} = E_{\max}/\omega_{\max} = 0,004775\;Tm^2$.

Based on the presented equations and data, we determine the electromechanical time constant of the motor in two variants

$$T_M = \frac{RJ}{0,5mk_{m1}^2} \text{ and } T_M = \frac{2RJ}{k_{m2}^2} \; . \qquad (16, 17)$$

Calculations of electromechanical time constants according to formulas (16, 17) gave the same result $T_M = 0,3842\,s$.

For further studies of the operating modes of the flywheel motor, we will use a simpler form of equations (10-15). However, it should be noted that when calculating the electromechanical time constant, it is necessary to take into account the active resistances of the transistors R_T of the voltage inverter and the current sensor R_I in the DC link, that is, the total resistance $R_\Sigma = 2R + 2R_T + R_I$ must be substituted into formula (17), then we get

$$T_M = \frac{\left(2R + 2R_T + R_I\right)J}{k_{m2}^2} \; . \qquad (18)$$

Taking into account the values $R_T = 0,045\,Ohm$ (for the IRLML2502 transistor) and $R_I = 0,05\,Ohm$ we have the total resistance $R_\Sigma = 0,766\,Ohm$ and the electromechanical time constant $T_M = 0,4701\,s$.

Based on equations (10-15), we write the transfer functions of the motor

$$\frac{I(p)}{U(p) - E(p)} = \frac{1/R_\Sigma}{T_E\,p + 1} \; ; \; \frac{\omega(p)}{\Delta M(p)} = \frac{1}{J\,p} \; ; \; \frac{\omega(p)}{U(p)} = \frac{1/k_{m2}}{T_M\,p + 1} \; . \qquad (19\text{-}21)$$

Let's supplement the description of the closed-loop system of automatic control of the flywheel motor angular speed with the controller transfer function $W_C(p)$, equations for determining the unbalance $U_U(p)$ and the feedback $U_{FB}(p)$ signals

$$\frac{U(p)}{U_U(p)} = W_C(p) \; ; \; U_U(p) = U_R(p) - U_{FB}(p) \; ; \; U_{FB}(p) = k_{FB}f(\omega) \; , \qquad (22\text{-}24)$$

where $U_R(p)$ is the rotor angular speed reference signal; k_{FB} is feedback gain; $f(\omega)$ is signal generator of the rotor angular speed in the form of a sequence of pulses with a reiteration period T_N.

We study the features of tuning the controller of the reactive flywheel motor angular speed control system, which operates under conditions of a relatively large value of the moment of inertia of the rotor, a slight mechanical load and a wide range of angular speed control.

The signal of the angular speed sensor corresponds to a sequence of N pulses per one revolution, while the area of each of the pulses is fixed over the entire range of the angular speed. In the simplest case, both duration and amplitude must be fixed. Let us determine the duration and amplitude of these pulses

$$t_1 = \frac{2\pi\,\gamma_{\max}}{N\,\omega_{\max}}; \quad U_{1\max} = \frac{\omega_{\max}}{\gamma_{\max}}, \quad (25, 26)$$

where $\gamma_{\max}$ is the maximum value of the duty cycle of the pulse sequence at the maximum value of the angular speed $\omega_{\max}$. The value of the duty cycle of the pulse sequence changes in proportion to the angular speed

$$\gamma = \frac{\omega\,\gamma_{\max}}{\omega_{\max}}. \quad (27)$$

And since the duty cycle γ can be in the range from 0 to 1, then for the calculation of the control system it is possible to recommend choosing the maximum value $\gamma_{\max}$ from the range of values from 0.5 to 0.8.

When designing the control system for the angular speed of the flywheel rotor, the pulse sequence of our sensor is converted into a continuous signal using a low-frequency filter, which simultaneously performs the function of an angular speed controller. Such a filter makes the structural and parametric stability of the system, an acceptable quality of regulation at an acceptable level of pulsation of the controller output signal over the entire range of angular speed. The functions of such a controller can be performed by an integrating link and a link with a combined transfer function

$$W_C(p) = \frac{k_C}{p} \quad \text{and} \quad W_C(p) = \frac{k_{C1}}{T_F\,p+1} + \frac{k_C}{p} = \frac{k_C\,(T_2\,p+1)}{p\,(T_F\,p+1)}, \quad (28, 29)$$

where k_C, k_{C1}, T_F, $T_2 = \dfrac{k_{C1}+T_F\,k_C}{k_C}$ are gains and time constants of the controllers.

Taking into account that the sequence of pulses of the angular speed sensor should be smoothed under the condition that the amplitude of pulsations at the output of the angular speed controller is limited to no more than a given value, for example, 10 % of the average signal value, then to calculate the parameters of controllers (28, 29) approaches of the theory of linear systems can be applied.

We write the transfer function of a closed astatic system (19-21) in the form

$$\frac{\omega(p)}{U_R(p)} = \frac{k_1}{T_1^2 p^2 + 2\xi T_1\,p + 1}, \quad (30)$$

where T_1, ξ are the time constant and damping factor of the second order link [8]. In this case, the parameters of the transfer function (30) are determined by the equalities

$$k_1 = \frac{1}{k_{FB}}; \quad T_1 = \frac{k_m T_M}{k_C\,k_{FB}}. \quad (31, 32)$$

For a stable oscillatory link, the condition $0 \le \xi \le 1$ is true. Thus, the tuning of the second-order control system (30) is determined only by the value of the damping factor ξ [8].

A feature of astatic control systems with controllers of the form (28, 29) is the equality to zero in the steady state of the average value of the unbalance signal $U_U(p)$ at the controller input. Therefore, the feedback gain of the astatic rotor angular speed control system is defined as

$$k_{FB} = \frac{U_{R\max}}{\omega_{\max}}, \quad (33)$$

where $U_{R\max}$ is the maximum value of the reference signal corresponding to the angular speed maximum value $\omega_{\max}$.

In [8], the relationships between the parameters of the control system, the pulse reiteration period T_N, the duty cycle γ and the relative magnitude ΔU^* of the signal ripples at the output of the angular speed controller were determined

$$\Delta U^* = \frac{u_{\max} - u_{\min}}{U_{av}} 100\%, \qquad (34)$$

where $u_{\max}$, $u_{\min}$, U_{av} are the maximum, minimum and average values of the controller output signal.

The studies described in [8] showed that two variants are possible when tuning a control system with an I-controller (28). In one case, it is possible to tune the system so that the specified value of the damping factor ξ is provided, and the relative value ΔU^* of the pulsations is less than the specified value. Or vice versa, when the specified relative value ΔU^* is reached, the value of the damping factor is greater than the specified value.

For a control system with an I-controller, the relationships between the open-loop gain k_{OLS} and the relative magnitude ΔU^* of ripples are determined, as well as between the relative magnitude ΔU^* of ripples and the damping factor ξ

$$k_{OLS} = \frac{N\,\omega_{\max}\,\omega\,\Delta U^*}{2\pi(\omega_{\max} - \omega\gamma_{\max})} \text{ and } \Delta U^* = \frac{\pi(\omega_{\max} - \omega\gamma_{\max})}{2N\xi^2 T_M \omega_{\max}\,\omega}. \qquad (35, 36)$$

In accordance with formula (35), depending on the given value ΔU^*, the open-loop gain k_{OLS} can be determined. And depending on the given value ξ, according to the formula (36), it is possible to calculate the relative value ΔU^* of pulsations. In this case, the gain of the I-controller is determined by the formulas

$$k_C = \frac{k_{OLS} k_m}{k_{FB}} \text{ and } k_C = \frac{k_m}{4\xi^2 T_M k_{FB}}. \qquad (37, 38)$$

Consider the conditions for tuning the control system with the I-controller. With the given parameters, using the formula (36), we determine the relative magnitude ΔU^* of the pulsations. If the calculated value ΔU^* is less than the specified value ΔU_R^*, then the calculation of the controller gain k_C is performed using formula (38). If the condition $\Delta U^* < \Delta U_R^*$ is not met, then the calculation of the controller gain k_C under the condition $\Delta U^* = \Delta U_R^*$ must be performed using formulas (35, 37).

Figure 2 shows the dependences of the relative magnitude of pulsations ΔU^*, expressed as a percentage, and the controller gain k_C (28) on the value of the motor angular speed in the lower part of the range of its operation up to 400 rpm. Here you can notice when there is a change in the fulfillment of the condition $\Delta U^* < \Delta U_R^*$. Calculations were performed with values ΔU_R^* equal to 10 (Fig. 2*a*) and 5 (Fig. 2*b*) percent, as well as with the number of feedback signal pulses per one revolution of the flywheel shaft $N=12$. In addition, here and below, other parameters of the system were also taken for calculations: $\xi = 0,7$, $U_{R\max} = 1$, $\gamma_{\max} = 0,5$, $M_L = 0$.

In contrast to an astatic system with an I-controller, in a system with a controller (29), to make the specified damping coefficient ξ and the relative magnitude ΔU^* of the controller output signal ripples, it is necessary to tune three parameters T_F, k_{C1} and k_{C2}. In accordance with [8], for a system with controller (29), the time constant T_F can be determined under the condition $T_2 = T_M$ as the result of solving the nonlinear equation

$$\Delta U^* = \frac{1}{4\xi^2 \gamma T_F}\left(\gamma(1-\gamma)T_N + (T_M - T_F)\left(1 - e^{-\frac{\gamma \cdot T_N}{T_F}}\right)\left(1 - e^{-\frac{(1-\gamma)T_N}{T_F}}\right)\left(1 - e^{-\frac{T_N}{T_F}}\right)^{-1}\right)100\%, \qquad (39)$$
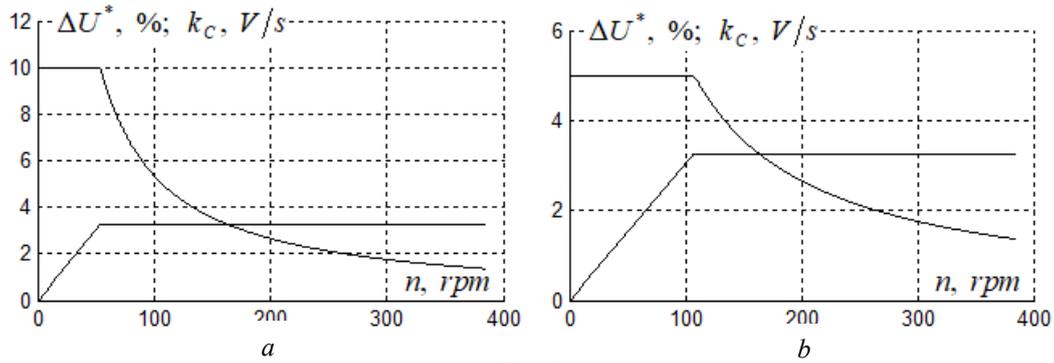
Fig. 2

where the parameters T_N, T_M and γ are determined by formulas (2, 18, 27). After obtaining the value of the time constant T_F, the remaining parameters of the controller (29) are determined as

$$k_C = \frac{k_m}{4\,\xi^2\,T_F\,k_{FB}} \quad \text{and} \quad k_{C1} = k_C\left(T_M - T_F\right). \tag{40, 41}$$

Under the condition $T_M < T_F$ it should be assumed $T_M = T_F$, then the angular speed controller becomes similar to the I-regulator.

Fig. 3 shows the dependences of the time constant on the magnitude of the motor angular speed, calculated at $\Delta U_R^* = 10\%$, $\xi = 0{,}7$ and *N*=12 in accordance with equation (39). For ease of perception of graphs, the dependence is divided into two subranges of angular speed, namely: from 0 to 300 rpm (Fig. 3*a*) and from 300 to 6000 (Fig. 3*b*). Fig. 4 shows the dependences of the controller gains k_C and k_{C1}, which are determined by formulas (40, 41) according to the time constant value T_F (Fig. 3).
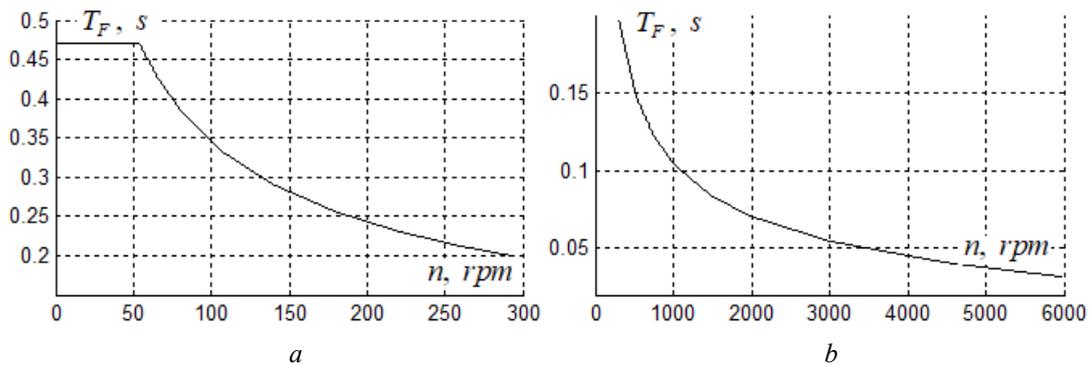
Fig. 3

Let us now study the influence of the number N of angular speed feedback signal pulses of the per one revolution of the flywheel shaft on the calculations of the time constant T_F of the controller filter. Fig. 5*a* shows two dependences of the time constant T_F on the number of pulses *N* at two values of the relative magnitude ΔU^* of the pulsations of 5 and 10 percent, respectively,
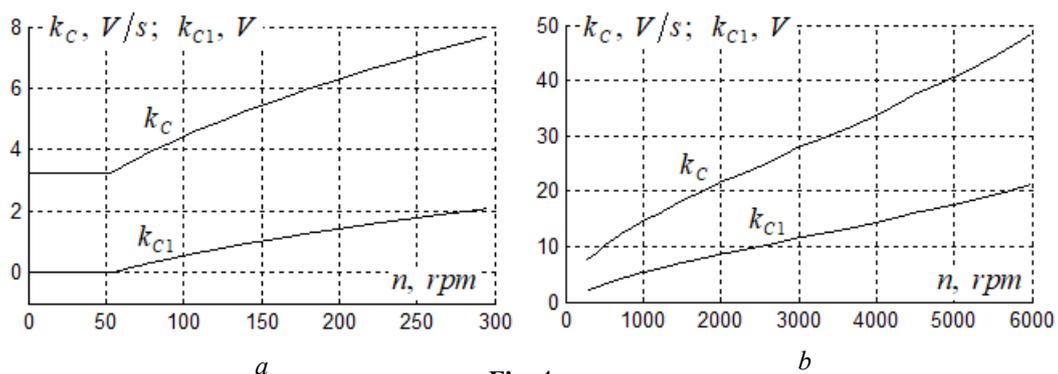
Fig. 4
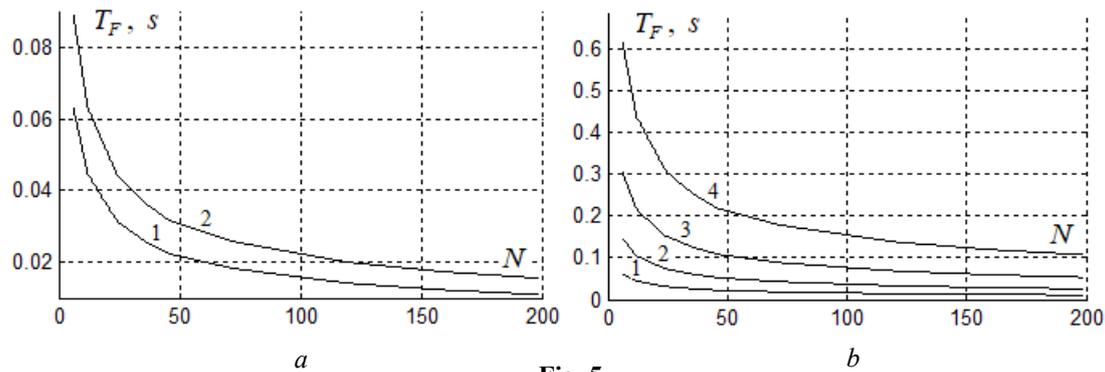
Fig. 5

indicated by the numbers 1 and 2. This calculation was performed at an angular speed of 4000 rpm for the range of change in the value of N from the minimum value 6 to 200. Fig. 5*b* shows four dependences of the time constant on the number of pulses N at $\Delta U^* = 10\%$ and four values of the angular speed of 62.5, 250, 1000 and 4000 rpm, respectively, indicated by the numbers 1, 2, 3 and 4.

The results of the calculation of transients are not presented in this paper, since, under the condition $\xi = 0,7$, the response of the second-order dynamic link (30) to a stepwise input action is characterized by an overshooting equal to 4.6 percent and a time to reach a five percent zone from the steady-state value of the output signal, which is approximately $4,1T_F$.

Conclusions. When tuning the speed controller, the traditional approach was used to compensate for the relatively large electromechanical time constant of the motor using the speed controller. In addition, the controller was tuned taking into account the pulsed nature of the angular speed sensor, provided that the given ripple range was obtained at the output of the speed controller. Such a study was carried out on the basis of the equation (39) obtained earlier by the authors for calculating the parameters of the regulator.

Studies have shown that the use of a first order filter with a time constant T_F in the speed controller to smooth out the ripple of the feedback signal limits the dynamic performance of the flywheel motor's angular speed control. Consideration of the dependences of the time constant T_F on the number of pulses N of the feedback signal on the angular speed showed their non-linear character. In this case, a significant effect of reducing the time constant T_F is achieved at $N=50$, and a further increase in the number of pulses N no longer leads to such a significant decrease in the time constant.

The approach to the formation of a feedback signal on the angular speed in the form of a sequence of pulses makes it possible to integrate such a mechanical coordinate sensor into the motor design without increasing its overall dimensions. In addition, this reduces the requirements for the accuracy of producing the sensor, since in the astatic control system, even with some violation of the uniformity of the pulse repetition period, a zero static error in processing the reference signal for angular speed is provided.

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НАЛАШТУВАННЯ РЕГУЛЯТОРА КУТОВОЇ ШВИДКОСТІ РЕАКТИВНОГО МАХОВИКА СИСТЕМИ ОРІЄНТАЦІЇ НАНОСУПУТНИКА

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У статті описано реактивні маховики на основі безконтактних магнітоелектричних двигунів для управління просторовою орієнтацією супутників, розроблені в Інституті електродинаміки НАН України. Основну увагу приділено проблемі реалізації роторного давача кутової швидкості та положення за умови його можливого вбудовування у конструкцію двигуна. Досліджено вплив кількості імпульсів за один оборот валу двигуна на налаштування регулятора кутової швидкості реактивного маховика. Отримано залежності сталої часу фільтра та коефіцієнтів передачі регулятора від величини кутової швидкості реактивного маховика та кількості імпульсів сигналу зворотного зв'язку за один оборот за умови обмеження розмаху пульсацій на виході регулятора на заданому рівні. Бібл. 8, рис. 5, таблиця.
Ключові слова: наносупутник, реактивний маховик, безконтактний магнітоелектричний двигун, система керування, давач кутової швидкості, регулятор кутової швидкості.